REMARKS:

Claims 1-21 are currently being examined, of which claims 13, 17, and 18 have been amended herein and claims 19-21 have been newly added herein.

Applicants and Applicants' attorney thank Examiner Leung for the interview courteously granted November 22, 2005. The special attention the Examiner paid to the instant application is noted with appreciation. Items discussed during the interview include the Office Action mailed July 25, 2005.

The Examiner has indicated that claims 13, 17, and 18 set forth allowable subject matter. In particular, the Examiner has objected to claims 13, 17, and 18 as being dependent upon rejected base claims, and has noted that claims 13, 17, and 18 would be allowable if rewritten in independent form including the limitations of the base claims and any intervening claims.

Claims 13, 17, and 18, as amended herein, are intended to correspond to previous claims 13/9/6/5/1, 17/9/6/5/1, and 18/9/6/5/1. Claims 19-21, as newly added herein, are intended to correspond to previous claims 13/9/7, 17/9/7, and 18/9/7.

Claims 13, 17, and 18 have been amended herein in a manner intended to place them in condition for allowance. Thus, in view of the above, Applicants respectfully submit that this

objection should be withdrawn. Accordingly, Applicants respectfully submit that claims 13 and 17-21 are in condition for allowance.

Claims 8, 10-12, and 16 stand rejected under the first paragraph of 35 USC 112 as failing to comply with the written description requirement. Claims 10-12 and 16 depend from claim 8. The Examiner has suggested that the subject application fails to describe in detail the following aspects of claim 8: "bandwidth of optical output of said Mach Zehnder light intensity modulator is restricted by using loss of said travelling wave type electrode."

It is well known to restrict bandwidth using a loss of an electrode. It is well known that loss of a radio frequency transmission line, such as coplanar waveguide, a microstrip line, and a strip line, depends upon frequency. The loss in a radio frequency transmission line includes the elements such as conductor loss, dielectric loss. The higher the frequency is, the higher the loss of all the elements of a radio frequency transmission line. In a Mach Zehnder type light modulator, high frequency component of optical output is decreased when loss of high frequency component in a travelling wave type electrode increases, and the optical bandwidth is restricted by the loss.

A document with a 1991 copyright is enclosed to show information relating to restricting bandwidth using a loss of an electrode, and to: (1) support the idea that it is well known to restrict bandwidth using a loss of an electrode; and (2) demonstrate that the rejection of claim 8 under the

first paragraph of 35 USC 112 should be withdrawn.

The enclosed document shows loss in a coplanar waveguide in equations 3.4.1.10 (dielectric loss), 3.4.1.11 (conductor loss), and 3.4.1.14 (radiation loss), which show that conductor loss is proportional to square root of frequency (f) (see Rs in 3.4.1.12), and dielectric loss and radiation loss are proportional to frequency (or inverse of wavelength λ_g). Thus, the desired loss characteristics are obtained by designing those losses and other parameters. As the loss increases as the frequency, the bandwidth is restricted by the loss.

In a Mach Zehnder type light modulator, high frequency component of optical output is decreased when loss of high frequency component in a travelling wave type electrode increases, and the optical bandwidth is restricted by the loss.

Thus, Applicants respectfully submit that the rejection of claim 8, 10-12, and 16 should be withdrawn.

Claims 1-5 stand rejected under 35 USC 102(b) as anticipated by USP 5,543,952 (**Yonenaga '952**) in reference to *Introduction to CMOS Design* (**Zitti**).

Claims 6, 7, 9, 14, and 15 stand rejected under 35 USC 103(a) as obvious over **Yonenaga**

in reference to **Zitti**.

Applicants respectfully traverse the above rejections of claims 1-7, 9, 14, and 15.

The Examiner suggests that **Yonenaga '952** (Fig. 1b, col. 3 at lines 33-45, col. 4 at lines 14-16, and col. 5, line 66 to col. 6, line 3) expressly or inherently describes all features set forth in claim 1, except the ***amplifier***. The Examiner relies on **Zitti** (page 7-11) to argue that an inverter is inherently an amplifier. Thus, the Examiner is suggesting that the inverter 11 of **Yonenaga '952** (see Fig. 1b) is inherently an amplifier.

However, the Examiner has not specifically identified a publication date of **Zitti** on Form PTO-892 or in the body of the Office action. Also, the Examiner has not established that the critical reference date of **Zitti** precedes the filing of the subject application. Thus, the Examiner has not yet demonstrated that **Zitti** may be used as prior art against claims of the subject application.

In response to the above rejections of independent claims 1 and 7, please note that:

(A) the inverter 11 of **Yonenaga '952** (Fig. 1b) does not describe, teach, or suggest the amplifier as set forth in claims 1 and 7;

(B) the Examiner has not identified a publication date of **Zitti** and has not established that

Zitti can be applied as prior art against claims of the subject application, and thus the rejections of claims 1 and 7 should be withdrawn; and

(C) even if **Zitti** can be applied as prior art against claims of the subject application, **Zitti** fails to remedy the deficiencies of **Yonenaga '952**.

Thus, Applicants respectfully submit that the rejections of claims 1 and 7, and all claims depending therefrom, should be withdrawn.

The Examiner has indicated that a certified copy of the priority document may not currently be in the appropriate file of the U.S. Patent and Trademark Office. However, a certified copy of the priority document was filed on August 28, 2001. A dated postcard receipt is enclosed to demonstrate that a certified copy of the priority document was filed on August 28, 2001.

In view of the aforementioned amendments and accompanying remarks, all claims currently being examined are in condition for allowance, which action, at an early date, is requested.

If, for any reason, it is felt that this application is not now in condition for allowance, the Examiner is requested to contact the Applicants' undersigned attorney at the telephone number indicated below to arrange for an interview to expedite the disposition of this case.

In the event that this paper is not timely filed, the Applicants respectfully petition for an appropriate extension of time. Please charge any fees for such an extension of time and any other fees which may be due now or in the future with respect to this application, to Deposit Account No. 01-2340.

Respectfully submitted,

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To all my teachers—especially my father and my beautiful wife, Andrea.

Contents

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[21] Young, Leo, ed., *Advances in Microwaves*, Volume 1, "Directional Couplers," by R. Levy, Academic Press, New York, 1966, pp. 115–209. (An excellent discussion.)

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Chapter 3

Physical Transmission Lines

3.1 INTRODUCTION

In this chapter we transform our knowledge about the generalized transmission line calculations into actual physical realizations for a wide variety of configurations. Structures are grouped into major categories: coaxial, paired lines, coplanar, microstrip line, stripline, waveguide, and others. Within each category, a number of nonstandard configurations are included to illustrate the effects of variations from the standard structure. Tables and graphs are generally excluded because it is assumed that the reader has access to and will use the *Transmission Line Design Software* available as part of this text.

ABCD matrices for the lines in this chapter are found in Chapter 2. The equations of this chapter allow calculation of Z_0 in the matrices from the line's physical dimensions. The electrical length is calculated from the operating frequency, f, and the physical length, l, using ε_{eff}.

$$\theta = 2.0\,\pi\, l f \,\frac{\sqrt{\varepsilon_{eff}}}{c} \;\text{(radians)}\; = 360.0\, l f\, \frac{\sqrt{\varepsilon_{eff}}}{c} \;\text{(degrees)} \qquad (3.1.1)$$

A number of the older and commonly referenced articles are available as part of reprint collections listed below. Two of these, [2] and [4] are out of print but they may be found in libraries. Consulting these can save considerable time.

REFERENCES

[1] Itoh, T., ed., *Planar Transmission Line Structures*, IEEE Press, NY, 1987.

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[3] Frey, J., and Kul Bhasin, eds., *Microwave Integrated Circuits*, 2nd Ed., Artech House, Norwood, MA, 1985. (The second edition of [2] with a new group of papers.)

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3.2 COAXIAL STRUCTURES

3.2.1 Round Coaxial Cable

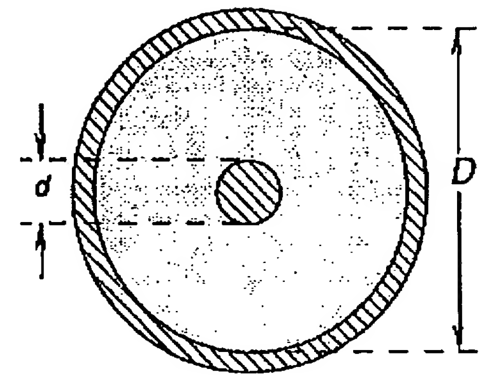

Figure 3.2.1.1: Round Coaxial Cable

The equations for this transmission line can be derived easily and are exact for the case of an infinitely long line (no fringing fields):

$$Z_0 = \frac{\eta_0}{2.0\ \pi \sqrt{\varepsilon_r}} \ln\left(\frac{D}{d}\right) \quad (\Omega) \qquad (3.2.1.1)$$

$$\alpha_c = \left(\frac{0.014272\sqrt{f}}{Z_0}\right)\left(\frac{1}{d} + \frac{1}{D}\right) \quad (\text{dB / m}) \qquad (3.2.1.2)$$

$$\alpha_d = 0.091207\, f \sqrt{\varepsilon_r} \tan\delta \quad (\text{dB / m}) \qquad (3.2.1.3)$$

For inner and outer conductors having equal conductivity, the optimal D / d for breakdown voltage is 2.718; for optimal power transfer, 1.65; and for minimum attenuation, 3.59 [7, 13]. See [7] for information on optimizing these parameters.

Coaxial connectors have mechanical tolerances of ±0.0025 mm and ±0.005 mm for laboratory and general precision, respectively. For a 50 Ω connector, this corresponds to about 0.5% and 0.8% tolerance on the impedance.

To reduce losses, we might increase d and D, keeping their ratio constant to lower α_c. However, the coaxial wire of Figure 3.2.1.1 will propagate higher order modes.

The cutoff frequency of these modes decreases with increasing d, and D placing a limitation on the cable size for a desired useful frequency range. Marcuvitz [11] gives the relations for the higher order modes, which begin at approximately 1.5 times the TE_{11} mode. The first higher order mode begins to propagate when

$$\lambda_{C_{11}} = \frac{2.0\,\pi}{\beta_{z1}} \tag{3.2.1.4}$$

where β_{zm} is the first ($m = 1$) solution of

$$\frac{J_1'(\beta_{zm} R)}{N_1'(\beta_{zm} R)} = \frac{J_1'(\beta_{zm} r)}{N_1'(\beta_{zm} r)} \tag{3.2.1.5}$$

where

$$r = d / 2.0 \tag{3.2.1.6}$$

$$R = D / 2.0 \tag{3.2.1.7}$$

Dimitrios [5] rewrites the above by replacing derivative forms of Bessel functions with nonderivative forms. This form is more accurate when used with available tables of Bessel functions:

$$\frac{J_0'(\beta_{zm} R) - \dfrac{1}{\beta_{zm} R} J_1'(\beta_{zm} R)}{N_0'(\beta_{zm} R) - \dfrac{1}{\beta_{zm} R} N_1'(\beta_{zm} R)} = \frac{J_0'(\beta_{zm} r) - \dfrac{1}{\beta_{zm} r} J_1'(\beta_{zm} r)}{N_0'(\beta_{zm} r) - \dfrac{1}{\beta_{zm} r} N_1'(\beta_{zm} r)} \tag{3.2.1.8}$$

Green [8] derives an expression for the wavelength at which this occurs (TE_{11} mode begins to propagate):

$$\lambda_c = 2\,\pi\, r_m \left[1.0 - \frac{1.0}{6.0} \left(\frac{t}{2.0\, r_m} \right)^2 - \frac{7.0}{120.0} \left(\frac{t}{2.0\, r_m} \right)^4 - \dots \right] \tag{3.2.1.9}$$

where

$$r_m = \frac{d + D}{2.0} \tag{3.2.1.10}$$

$$t = D - d \tag{3.2.1.11}$$

The stated agreement with Marcuvitz's Bessel function solution is better than 1% for $D / d < 2.5$, and better than 4% for $D / d < 4.0$. The number of terms used to reach this agreement is not stated.

A simple approximation to the above is:

$$\lambda_c \cong \frac{\pi\,(D + d)}{2.0} \quad \text{(units of } D, d\text{)} \tag{3.2.1.12}$$

which is the circumference of a circle with radius midway between the inner and outer conductors. Dimitrios [5] compared the accuracy of the equation above to the exact solution and found that it is generally accurate to 3% for 50 Ω lines with various dielectrics.

A lesser known fact about real-world coaxial cables is that they can introduce nonlinear distortion at a low level. Amin [3] presents data at the L, S, and C (390 MHz to 10.9 GHz) bands for various commercial cables showing that braided cables can have distortion products down 90 to 115 dB from the carrier depending on construction. Solid shields had no distortion. Nickel-plated, stainless steel, and Al alloy braids had distortion down 90 to 95 dB from the carrier. Distortion increased with power, frequency, and cable length.

3.2.1.1 *Example: Coaxial Pogo Pin*

We wish to design a 50 Ω controlled impedance path that launches from a microstrip line and makes use of a standard size (0.040" o.d.) pogo pin (a spring-loaded pin used in ATE equipment) as the center conductor. This configuration is of use in test equipment where a controlled Z path must be blind-mated repetitively, as in Figure 3.2.1.2. Calculate the required shield dimensions for both air and PTFE dielectrics.

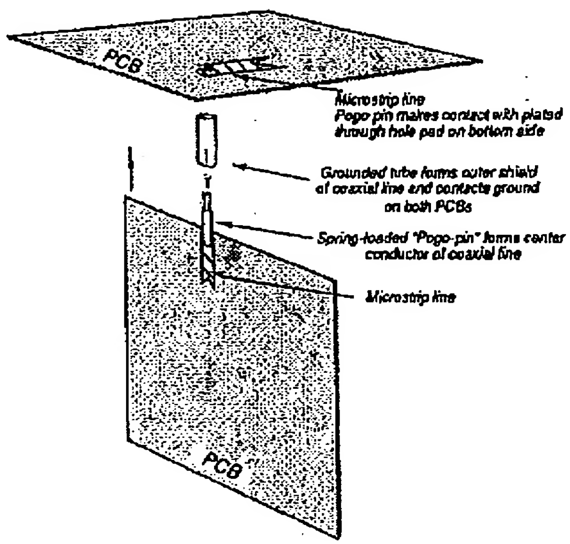

Figure 3.2.1.2: Coaxial Pogo-Pin Example

Solution:

$$Z_0 = 50\ \Omega = \frac{\eta_0}{2.0\ \pi \sqrt{\epsilon_r}} \ln\left(\frac{D}{d}\right)\Omega$$

For air:

$$50\ \Omega = \frac{377}{2\ \pi\sqrt{1}} \ln\left(\frac{D}{0.040}\right)\Omega$$

$$\boxed{D = 0.092"}$$

For PTFE:

$$50\ \Omega = \frac{377}{2\ \pi\sqrt{2.1}} \ln\left(\frac{D}{0.040}\right)\Omega$$

$$\boxed{D = 0.134"}$$

A number of companies have attempted to combine the 360° shielding benefits of the coaxial structure with the fabrication advantages of printed-circuit boards (PCBs). This is increasingly an issue as the density and speeds of digital circuitry increase. See also [15] for a planar structure constructed with additive techniques. See Section 3.2.5 for more information.

Swengel *et al.* [18] describe a technique for machine wiring coaxial lines on PCB. The technique point-to-point wires 3.14 mil wires coated with a 10 mil thick layer of PTFE. The wire ends are then plated together. The PTFE is plated, creating the coaxial shield. A final selective etch separates the wire ends as required. Because of the solid ground plane, wave-soldering is difficult. The technique is also sensitive to handling and vibration.

A similar technique, described in [21], uses a semirigid cable (PTFE, $D = 9.5$ mil, $d = 3.14$ mil), which is also machine routed point-to-point. The wires are laid down on an adhesive layer during routing and then the shields are plated together. The wire is stripped at its ends, coated with epoxy and the center conductors are plated to plated-through holes. This structure is easily used as part of a multilayer PCB.

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3.2.2 Partially Filled Round Coaxial Cable

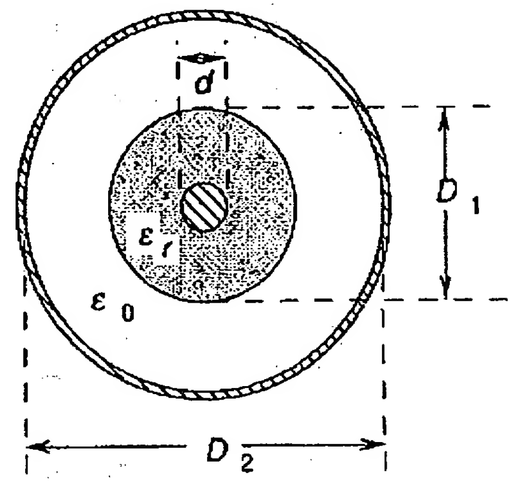

Figure 3.2.2.1: Partially Filled Round Coax

This case is encountered where a low-loss line is being constructed and the dielectric losses need to be reduced by making part of the dielectric air. It also occurs when the dielectric diameter is reduced over a portion of its length for assembly.

$$Z_0 = \frac{\eta_0}{2\pi} \ln\left(\frac{D_2}{d}\right) \sqrt{\frac{\varepsilon_r \ln\left(\frac{D_2}{d}\right) + \ln\left(\frac{D_1}{d}\right)}{\varepsilon_r \ln\left(\frac{D_2}{d}\right)}} \quad (\Omega) \qquad (3.2.2)$$

Ragan [3] also discusses the matching of such structures by proper choice of dimensions. Hatsuda [1] analyzes a round coaxial structure with wedges of dielectric removed.

3.2.2.1 *Example: Calculating q for a Partially Filled Coaxial Line*

Solution: The ratio of the structure's actual capacitance to its capacitance with solid air as the dielectric is defined as q. We can easily get C_{air} from the solid dielectric structure:

$$C_{air} = \frac{2\,\pi\,\varepsilon_0}{\ln\left(\frac{D_2}{d}\right)}$$

and using the relationships in Chapter 1 we can write the capacitance in terms of the known Z_0:

$$C_{actual} = \frac{\sqrt{\mu_0\,\varepsilon}}{Z_0}$$

$$q = \frac{C_{actual}}{C_{air}} = \frac{\frac{\sqrt{\mu_0\,\varepsilon}}{Z_0}}{\frac{2\,\pi\,\varepsilon_0}{\ln\left(\frac{D_2}{d}\right)}} = \frac{\frac{\sqrt{\mu_0\,\varepsilon}}{\frac{\eta_0}{2\,\pi}\ln\left(\frac{D_2}{d}\right)\sqrt{\frac{\varepsilon_r\ln\left(\frac{D_2}{d}\right)+\ln\left(\frac{D_1}{d}\right)}{\varepsilon_r\ln\left(\frac{D_2}{d}\right)}}}}{\frac{2\,\pi\,\varepsilon_0}{\ln\left(\frac{D_2}{d}\right)}}$$

$$\boxed{q = \frac{\varepsilon_r}{\sqrt{\frac{\varepsilon_r\ln\left(\frac{D_2}{d}\right)+\ln\left(\frac{D_1}{d}\right)}{\varepsilon_r\ln\left(\frac{D_2}{d}\right)}}}}$$

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3.2.3 Eccentric Round Coaxial Cable

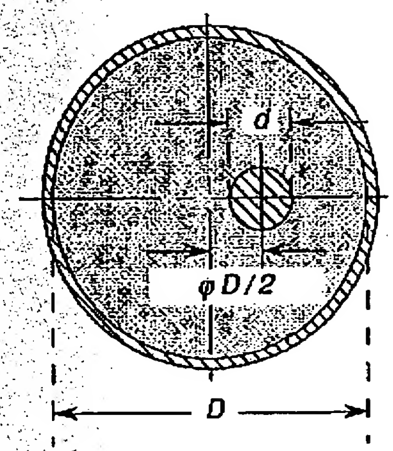

Figure 3.2.3.1: Eccentric Round Coaxial Cable

The eccentric coax equations enable us to analyze the effects of tolerances in the manufacture of the cable. This structure has also been used as a continuously adjustable $\lambda/4$ line. The center conductor is moved within the outer shield by a mechanical probe resulting in smooth variation of the characteristic impedance.

For center conductors off center in one direction, [1] gives:

$$Z_0 = \frac{\eta_0}{2.0\,\pi\sqrt{\varepsilon_r}}\cosh^{-1}\left[\frac{D}{2.0\,d}\left(1.0-\varphi^2\right)+\frac{d}{2.0\,D}\right](\Omega) \tag{3.2.3.1}$$

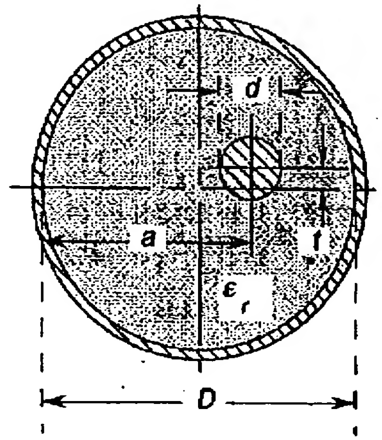

Figure 3.2.3.2: Eccentric Round Coaxial Cable

For center conductors that are eccentric in both directions, as in Figure 3.2.3.2 [2]:

$$Z_0 = \left[\frac{\eta_0}{2\pi\sqrt{\varepsilon_r}} \ln\left(\frac{D}{d}\right)\right]\left[1.0 - \frac{e^2 k^2}{(k^2 - 1.0)\log k}\right] \ (\Omega) \qquad (3.2.3.2)$$

where

$$b = \frac{d}{2.0} \qquad (3.2.3.3)$$

$$c = t \qquad (3.2.3.4)$$

$$e = \frac{c}{a} \qquad (3.2.3.5)$$

$$k = \frac{a}{b} \qquad (3.2.3.6)$$

The conductor losses of this line are

$$\alpha_c = \alpha_{c,centered}\left[1.0 + \frac{2.0\,e^2}{k} - \frac{e^2 k^2}{(k^2 - 1.0)\log k}\right] \quad (\mathrm{dB/m}) \qquad (3.2.3.7)$$

$\alpha_{c,centered}$ = conductor losses of the centered line calculated with (3.2.1.2).

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3.2.4 Square Coaxial Cable

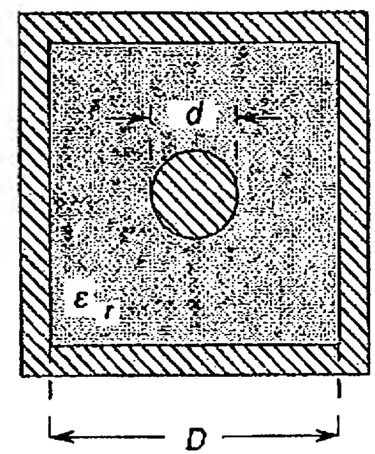

Figure 3.2.4.1: Square Coaxial With Circular Center Conductor

The square coaxial wire configurations are useful at very high frequencies as smaller replacements for waveguide. The square cross section makes them easier to fabricate than a round coax.

For the round center conductor as shown in Figure 3.2.4.1:

$$Z_0 = \frac{\eta_0}{2\pi\sqrt{\varepsilon_r}} \ln\left[\frac{1.0787\,D}{d}\right] \ (\Omega) \qquad (3.2.4.1)$$

and for $Z_0 \le 2.0\,\Omega$

$$Z_0 = 21.2\sqrt{D/d - 1.0} \quad (\Omega) \qquad (3.2.4.2)$$

which was derived by S. Frankel. Error of the first equation is less than 1.5% above 17 Ω and very small above 30 Ω [2]. The second equation is accurate to 0.5 Ω.

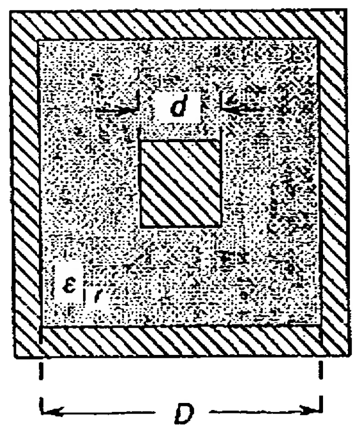

Figure 3.2.4.2: Square Coax with Square Center Conductor

For the square center conductor configuration shown in Figure 3.2.4.2 a conformal mapping solution is available [6]:

$$Z_0 = 5.0\,\pi\,v_p \times 10^{-8}\frac{K(k')}{K(k)} = \frac{5.0\,\pi\,c}{\sqrt{\varepsilon_r}}\frac{K(k')}{K(k)} \times 10^{-8} \quad (\Omega) \qquad (3.2.4.3)$$

$$k = \frac{(\Lambda' - \Lambda)^2}{(\Lambda' + \Lambda)^2} \qquad (3.2.4.4)$$

$$\Lambda' = \sqrt{1.0 - \Lambda^2} \qquad (3.2.4.5)$$

where Λ is the solution to:

$$\frac{K(\Lambda')}{K(\Lambda)} = \frac{D-d}{D+d} \qquad (3.2.4.6)$$

This can be solved iteratively with an equation solver or using the relations of Chapter 12. Observe that the above is always ≥0 and ≤1.

An approximation [6] is:

$$Z_0 \equiv \frac{1.0}{4.0\,v_p\,(C_{pp} + C_c)} = \frac{\eta_0}{\sqrt{\varepsilon_r}}\left[\frac{1.0}{4.0\left(\frac{2.0\,d}{D-d} + 0.558\right)}\right] (\Omega) \qquad (3.2.4.7)$$

which can be seen to be the equation for four parallel plate capacitors (C_{pp}) added to the four corner capacitances (C_c). The accuracy of the above equation is better than 1% for

$D/d \leq 4.0$

REFERENCES

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3.2.5 Rectangular Coaxial Line

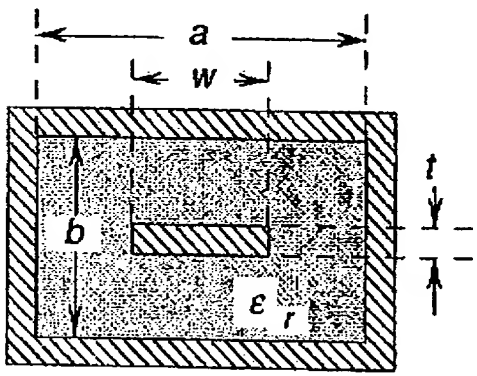

Figure 3.2.5.1: Rectangular Coaxial Line

This structure is a transitional structure between round coax and stripline, microstrip line or other planar lines. Many attempts have been made to utilize this structure for photolithographic construction of fully shielded controlled impedance lines. Rotating the center conductor relative to the shield also allows this structure to have a continuously variable characteristic impedance [4].

As reported in [9] rectangular coax can be used in PCBs constructed with additive techniques to obtain a completely shielded structure. The process by Augat Microtec builds the structure using photolithographic techniques from polyimide and copper.

A related structure [14], Figure 3.2.5.2, was constructed with thick-film techniques. The main limitation appears to have been the thickness, b, achievable without excessive layered printings. Without a sufficiently thick dielectric, center conductor to shield shorts were common. Low dielectric constant pastes were recommended.

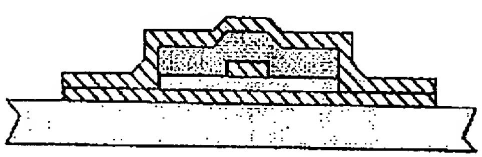

Figure 3.2.5.2: Thick-Film Coaxial Line

$$Z_0 = \frac{\eta_0}{4.0\sqrt{\epsilon_r}}\left[\frac{1.0}{\frac{w/b}{1.0 - t/b} + \frac{2.0}{\pi}\ln\left(\frac{1.0}{1.0 - t/b} + \coth\frac{\pi a}{2.0\,b}\right)}\right](\Omega) \quad (3.2.5.1)$$

for a square shield and center conductor, $a = b$ and $w = t$:

$$Z_0 = \frac{\eta_0}{4.0\sqrt{\epsilon_r}}\left[\frac{1.0}{\frac{1.0}{b/t - 1.0} + \frac{2.0}{\pi}\ln\left(\frac{1.0}{1.0 - t/b} + \coth\frac{\pi}{2.0}\right)}\right](\Omega) \quad (3.2.5.2)$$

As stated earlier, rectangular coaxial structures are also useful at extremely high frequencies as a smaller replacement for waveguide that is more easily fabricated than round coax.

Rotation of the center conductor varies the characteristic impedance of the line. The equations are in Cruz and Brooke, [4]:

$$Z(\theta) = \frac{Z(0°)\,(1.0 + \cos 2.0\theta) + Z(90°)\,(1.0 - \cos 2.0\theta)}{2.0}\ (\Omega) \quad (3.2.5.3)$$

where $Z(0°)$ and $Z(90°)$ are calculated with any of the equations for a rectangular line above. Although the calculations of $Z(0°)$ and $Z(90°)$ were out of range for the [2] and [1] equations, agreement with experimental data to better than 1.4% was achieved.

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"Characteristic Impedance of Rectangular Coaxial Transmission Lines,"" Vol. MTT-32, No. 2, February 1984, p. 219.

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3.2.6 Trough Line or Channel Line

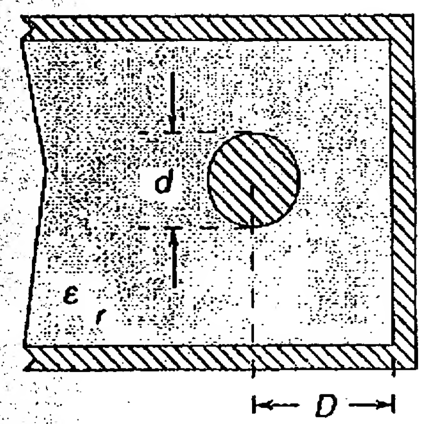

Figure 3.2.6: Trough Line

Wheeler [2] gives:

$$Z_0 = \frac{\eta_0}{4\pi\sqrt{\varepsilon_r}} \ln\left[1.0 + \left\{\frac{1}{2}\left(\frac{4}{\pi}\tanh\frac{\pi}{2}\right)^2\left[(D/d)^2 - 1.0\right]\right\} + \sqrt{\left\{\frac{1}{2}\left(\frac{4}{\pi}\tanh\frac{\pi}{2}\right)^2\left[(D/d)^2 - 1.0\right]\right\}^2 + \frac{4}{9}\left[(D/d)^2 - 1.0\right]}\,\right] \quad (\Omega) \qquad (3.2.6)$$

with a stated accuracy of about 1%.

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3.2.7 Strip-Centered Coaxial Line

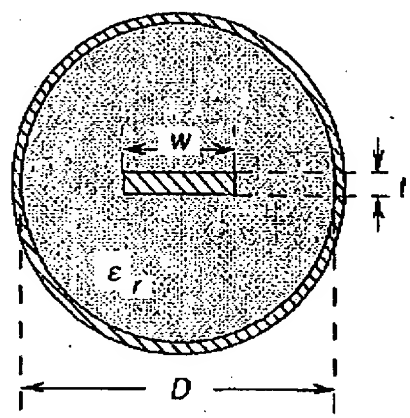

Figure 3.2.7.1: Strip-Centered Coaxial Cable

The configuration above is used to obtain lower loss by building a coaxial cable with a larger center conductor surface area. Bongianni [1] reports cables made with extremely fine dimensions—the center conductor was 150 µm × 12.5 µm. For $Z_0 \le 30.0\,\pi / \sqrt{\varepsilon_r}$ (Ω):

$$Z_0 = \frac{15.0\,\pi^2}{\sqrt{\varepsilon_r}} \frac{1.0}{\ln\left[\frac{2.0(D+w)}{D-w}\right]} \quad (\Omega) \tag{3.2.7.1}$$

For $Z_0 \ge 30.0\,\pi / \sqrt{\varepsilon_r}$ (Ω):

$$Z_0 = \frac{60.0}{\sqrt{\varepsilon_r}} \ln\left(\frac{2.0\,D}{w}\right) \quad (\Omega) \tag{3.2.7.2}$$

There is a bit of a chicken-and-the-egg problem here because it is necessary to know Z_0 in order to choose the equation for Z_0; however, the two equations pass quite close and it is okay to use either to make the choice.

REFERENCES

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3.3 PAIRED LINES

3.3.1 Parallel Wires

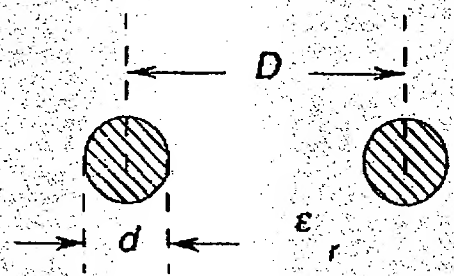

Figure 3.3.1.1: Parallel Wires

$$Z_0 = \frac{\eta_0}{\pi\sqrt{\varepsilon_r}} \cosh^{-1}\left(\frac{D}{d}\right) \ (\Omega) \qquad (3.3.1.1)$$

or

$$Z_0 = \frac{\eta_0}{2.0\,\pi\sqrt{\varepsilon_r}} \cosh^{-1}\left(\frac{2.0\,D^2 - d^2}{d^2}\right) \ (\Omega) \qquad (3.3.1.2)$$

valid for

$d/D << 1.0$

The first reference is found in [4] and [5]. The second was derived with conformal mapping techniques in [2] with a stated accuracy of better than 0.24%. Green, *et al.* [1] gives the conductor losses as:

$$\alpha_c = \frac{P}{2.0\,d} \frac{\sqrt{\frac{f\,\varepsilon}{\sigma}}}{\cosh^{-1} v} \quad \text{(nepers / cm)} \qquad (3.3.1.3)$$

where

$$P = \frac{v}{\sqrt{v^2 - 1.0}} \qquad (3.3.1.4)$$

$$v = \frac{D}{d} \qquad (3.3.1.5)$$

and P is a proximity factor; the equation is valid for high frequencies.

REFERENCES

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3.3.2 Unequal Size Parallel Wires

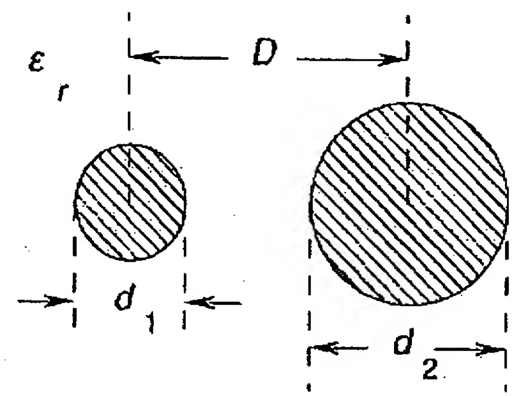

Figure 3.3.1.1: Unequal Size Parallel Wires

η_0 ... $(4.0\,D^2 - d_1^2 - d_2^2$ ...

The stated accuracy [1] is better than 0.24%.

REFERENCES

[1] Hilberg, Wolfgang, *Electrical Characteristics of Transmission Lines*, Artech House, Norwood, MA, 1979.

3.3.3 Twisted Pair

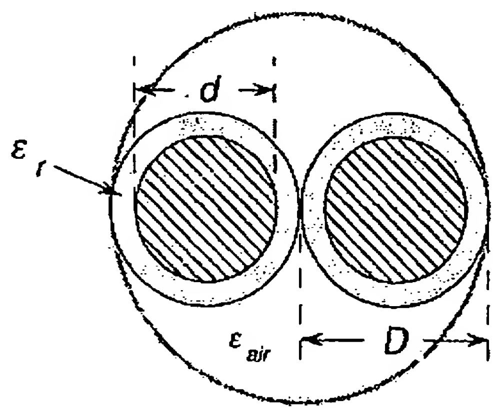

Figure 3.3.3.1: Twisted Pair

The twisted pair provides good low frequency shielding. Undesired signals tend to be coupled equally into each line of the pair. A differential receiver will therefore completely cancel the interference. Lefferson [3] gives design equations for this configuration:

$$Z_0 = \frac{\eta_0}{\pi \sqrt{\varepsilon_{eff}}} \cosh^{-1}\left(\frac{D}{d}\right) \quad (\Omega) \tag{3.3.3.1}$$

$$\varepsilon_{eff} = 1.0 + q\,(\varepsilon_r - 1.0) \tag{3.3.3.2}$$

$$q = 0.25 + 0.0004\,\theta^2 \tag{3.3.3.3}$$

$$T = \frac{\tan\theta}{\pi D} = \text{twists per length} \tag{3.3.3.4}$$

or

$$\theta = \tan^{-1}(T\,\pi D) \tag{3.3.3.5}$$

where T and D have the same length units and θ is the pitch angle of the twist; the angle between the twisted pair's center line and the twist. It was found to be optimal for θ to be between 20 and 45°. Smaller angles make the twist loose and create problems in maintaining tolerances. Angles above approximately 50.5° break the wire. The value of q was determined by fitting a line to measurements of the effective dielectic constant.

For the softer insulation PTFE, a different equation should be used for q

$$q = 0.25 + 0.001\,\theta^2 \tag{3.3.3.6}$$

An equation for the wire's total length before twisting in terms of number of turns, N, is:

$$l = N\,\pi D \sqrt{1.0 + \frac{1.0}{\tan^2\theta}} \tag{3.3.3.7}$$

Lefferson found these equations sufficiently accurate to design transmission lines with VSWR ≤ 1.1:1.

In [2] a new analysis and experimental data are presented. The derived equations are

$$Z_0 = \sqrt{\frac{L}{C}} \quad (\Omega) \tag{3.3.3.8}$$

$$\varepsilon_{eff} = \frac{C}{C_{air}}$$

where

$$L = \left(\frac{\mu_0}{\pi}\right)\cosh^{-1}\left(\frac{D}{d}\right) \tag{3.3.3.9}$$

$$C = C_1 + C_2 - C_3 \tag{3.3.3.10}$$

$$C_1 = \int_a^b \frac{\varepsilon_0\,dx}{D + (1.0/\varepsilon_r - 1.0)\sqrt{D^2 - x^2} - \dfrac{\sqrt{d^2 - x^2}}{\varepsilon_r}} \tag{3.3.3.11}$$

$$C_2 = \frac{\pi\,\varepsilon_0}{\ldots} \tag{3.3.3.12}$$

$$C_3 = \int_a^b \frac{\varepsilon_0 \, dx}{D - \sqrt{d^2 - x^2}} \qquad (3.3.3.13)$$

Calculate C_{air} by replacing ε_r with ε_0 in the equations for C. The integrals are evaluated by the Romberg numerical technique. Accuracy was tested by comparison with measurements of the line impedance. The technique was found to be accurate within the tolerances of the measurement and the film thickness variations.

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3.3.4 Five-Wire Line

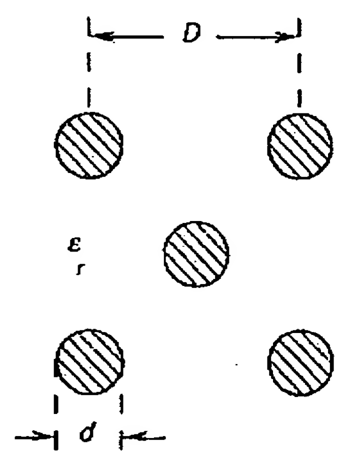

Figure 3.3.4.1: Five-Wire Line

$$Z_0 = \frac{2.5044\,\eta_0}{\ldots} \ln\left[\frac{D}{\ldots}\right] \quad (\Omega) \qquad (3.3.4)$$

where

$d/D \ll 1.0$

Of course, five conductors means lots of modes, so although the reference doesn't state it an assumption is that $(D - d)$ is small relative to a wavelength.

REFERENCES

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3.3.5 Paired Strips

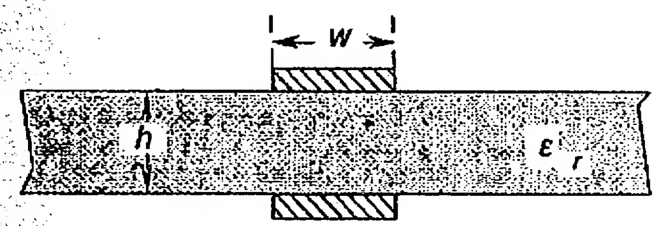

Figure 3.3.5.1: Paired Strips

For wide strips $(a/b > 1)$:

$$Z_0 = \frac{\eta_0}{\sqrt{\varepsilon_r}} \left\{ \frac{a}{b} + \frac{1.0}{\pi} \ln 4 + \frac{\varepsilon_r + 1.0}{2\,\pi\,\varepsilon_r} \ln\left[\frac{\pi\,\varepsilon\,(a/b + 0.94)}{2.0}\right] + \frac{\varepsilon_r - 1.0}{2\,\pi\,\varepsilon_r^2} \ln \frac{\varepsilon\,\pi^2}{16.0} \right\}^{-1} \quad (\Omega) \qquad (3.3.5.1)$$

Stated error is less than 1% for wide strips.

For narrow strips $(a/b < 1)$:

$$Z_0 = \frac{\eta_0}{\pi\sqrt{\varepsilon_r}} \left[\ln \frac{4.0\,b}{a} + \frac{1.0}{8.0}\left(\frac{a}{b}\right)^2 - \frac{\varepsilon_r - 1.0}{2.0\,(\varepsilon_r + 1.0)} \left(\ln \frac{\pi}{2.0} + \frac{\ln \frac{4.0}{\pi}}{\varepsilon_r} \right) \right] \quad (\Omega) \qquad (3.3.5.2)$$

where

$$b = h / 2.0 \qquad (3.3.5.4)$$

Stated error is less than 1% for narrow strips.

Equations are valid for $2b$ much smaller than half a wavelength in the dielectric (λ_g).

REFERENCES

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3.4 COPLANAR STRUCTURES

3.4.1 Coplanar Waveguide

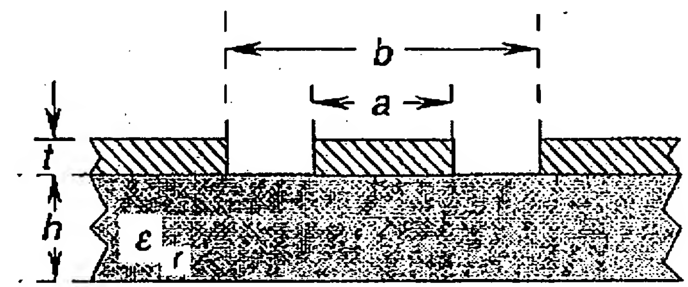

Figure 3.4.1.1: Coplanar Waveguide

The coplanar waveguide (CPW) configuration's advantages stem from its single-sided nature. Grounding components does not require plated through-holes to a plane on the other side of the substrate. This makes it ideal for use with surface mounted components. Another feature of coplanar waveguide is that we can narrow traces to match component lead widths while keeping Z_0 constant. The ground plane should extend greater than $5b$ on each side of the gap or the coplanar strip analysis should be used. The ground planes on either side of the center conductor may need to be connected periodically with wire jumpers depending on the frequency. The enclosure's cover can be used to jumper the two ground planes if it is kept away by at least $(b - a)$. To prevent propagation of higher modes, b should be less than $\lambda / 2$.

The design equations for coplanar waveguide are:

$$Z_0 = \frac{30.0\,\pi}{\sqrt{\varepsilon_{eff,t}}} \frac{K(k_t')}{K(k_t)} \qquad (3.4.1.1)$$

$$\varepsilon_{eff,t} = \varepsilon_{eff} - \frac{\varepsilon_{eff} - 1.0}{\frac{(b - a)/2.0}{0.7\,t} \frac{K(k)}{K'(k)} + 1.0} \qquad (3.4.1.2)$$

$$\varepsilon_{eff} = 1.0 + \frac{\varepsilon_r - 1.0}{2.0} \frac{K(k')K(k_1)}{K(k)K(k_1')} \qquad (3.4.1.3)$$

$$k_t = \frac{a_t}{b_t} \qquad k = \frac{a}{b} \tag{3.4.1.4}$$

$$k_t' = \sqrt{1.0 - k_t^2} \qquad k' = \sqrt{1.0 - k^2} \tag{3.4.1.5}$$

$$k_1 = \frac{\sinh\left(\frac{\pi a_t}{4.0\, h}\right)}{\sinh\left(\frac{\pi b_t}{4.0\, h}\right)} \tag{3.4.1.6}$$

$$k_1' = \sqrt{1.0 - k_1^2} \tag{3.4.1.7}$$

$$a_t = a + \frac{1.25\, t}{\pi}\left[1.0 + \ln\left(\frac{4.0\, \pi\, a}{t}\right)\right] \tag{3.4.1.8}$$

$$b_t = b - \frac{1.25\, t}{\pi}\left[1.0 + \ln\left(\frac{4.0\, \pi\, a}{t}\right)\right] \tag{3.4.1.9}$$

These equations are corrected for thickness which is important for accurate results. The dielectric losses are described in [6, 11] by:

$$\alpha_d = \frac{q\, \varepsilon_r \tan\delta}{\varepsilon_{eff}\, \lambda_g} \quad (\text{Np}/\text{m}) \tag{3.4.1.10}$$

$\frac{1}{f} \propto \frac{1}{\lambda_g}$

where q and λ_g have been previously defined.

The conductor losses are solved numerically and plotted in [6]. Jackson [9] reports that in some cases coplanar waveguide can have lower losses and dispersion than microstrip line for impedances near 50 Ω.

An equation for conductor losses [5] is:

$$\alpha_c = \frac{R_s \sqrt{\varepsilon_{eff}}\,[\Phi(a) + \Phi(b)]}{480.0\, \pi\, K(k)\, K(k')\, k'} \quad (\text{Np}/\text{m}) \tag{3.4.1.11}$$

where

$$R_s = \text{surface resistivity} = \sqrt{\pi f \mu_0 / \sigma} \quad (\Omega/\square) \tag{3.4.1.12}$$

$$\Phi(x) = \frac{\pi}{\ldots} + \frac{\ln\left[\frac{8.0\, \pi\, x\, (1.0 - k)}{t\, (1.0 + k)}\right]}{\ldots} \tag{3.4.1.13}$$

These equations are valid for

$t \ll a$

$t \gg b - a$

Radiation losses are

$$\alpha_r = \frac{\pi}{Q_{rad}\, \lambda_g} \tag{3.4.1.14}$$

$$Q_{rad} = \frac{K(k)\, K(k')}{\Psi(\varepsilon_r, h, k_0, b)\, \psi_{sc}\, \psi_{oc}} \tag{3.4.1.15}$$

$$\begin{aligned}\Psi(\varepsilon_r, h, k_0, b) &= (\varepsilon_r - 1.0)\left[1.0 + 0.5\, (\varepsilon_r - 1.0)^2 (k_0 h)^2\right] \\ &\times (k_0 h)(k_0 b)^2 \left[1.0 + 0.25\, (\varepsilon_r - 1.0)^2 (k_0 h)^2\right] \\ &\times \left\{\frac{1.0}{\sqrt{\varepsilon_{eff}}\, [1.0 + (\varepsilon_r - 1.0)^2 (k_0 h)^2]}\right\}\end{aligned} \tag{3.4.1.16}$$

$$\begin{aligned}\psi_{sc} &= \frac{5.0\, \pi}{8}\, \frac{1.0 + \left(1.0 - \frac{\pi^2}{8.0}\right)\left[1.0 + (\varepsilon_r - 1.0)^2 (k_0 h)^2 / 4.0\right]}{3.0\, \varepsilon_{eff}} \\ &\times \frac{\left[1.0 + (\varepsilon_r - 1.0)^2 (k_0 h)^2 / 4.0\right]}{\varepsilon_{eff}}\end{aligned} \tag{3.4.1.17}$$

$$\psi_{oc} = \frac{\pi}{8}\, \frac{3.0 + \left(1.0 - \frac{\pi^2}{8.0}\right)\left[1.0 + (\varepsilon_r - 1.0)^2 (k_0 h)^2 / 4.0\right]}{\varepsilon_{eff}} \tag{3.4.1.18}$$

These are valid for thin dielectric.

In Houdart [8, Figure 2], the effect of ground plane width (c in Figure 3.4.8.1) on the CPW is analyzed. For $c/b > 5.0$ the impedance is affected by less than about 3%.

3.4.1.1 *Example: Hybrid IC Probe in Coplanar Waveguide*

What are the required dimensions for a 50 Ω coplanar waveguide probe to a 10 mil ×

Solution: Assuming the probe is the same material as the hybrid IC substrate and using the above equations with

$h = 0.0252$ in $= 0.0640$ cm

$a = 0.010$ in $= 0.0254$ cm

$t = 0.001$ in $= 0.003$ cm

$\varepsilon_r = 10.0$

The program iteratively adjusts b until the goal impedance 50 Ω is achieved. The final dimension of the probe is:

$\boxed{b = 0.0234 \text{ in.} = 0.615 \text{ cm}}$

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3.4.2 Micro-Coplanar Stripline

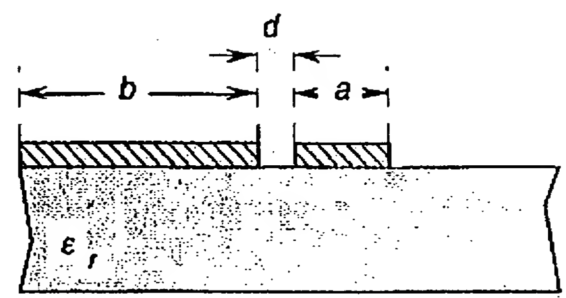

Figure 3.4.2.1: Micro-Coplanar Stripline

This structure has been called nonsymmetrical coplanar waveguide; however, to avoid confusion we prefer to reserve this term for coplanar waveguide having unequal gaps.

Kneppo and Gotzman [2] proposed this structure as an improvement over CPW for the connection of shunt elements. Their equations were derived with conformal transformation:

$$Z_0 = \frac{\eta_0}{\left(\sqrt{\varepsilon_1} + \sqrt{\varepsilon_2}\right)} \frac{K(k)}{K'(k)} \quad (\Omega) \tag{3.4.2.1}$$

$$\varepsilon_{eff} = \frac{\left(\sqrt{\varepsilon_1} + \sqrt{\varepsilon_2}\right)^2}{4.0} \tag{3.4.2.2}$$

where

$$k = \sqrt{\frac{1.0 + a/d + b/d}{(1.0 + b/d)\,(1.0 + a/d)}} \tag{3.4.2.3}$$

Polynomial equations for the line width, ε_{eff}, and conductor losses are given in Quian and Yamashita [3] for dielectrics with $\varepsilon_r = 2.22$, 9.7, 10.1, and 12.9 and in Yamashita *et al.* [4] for $\varepsilon_r = 12.7$.

Losses may be calculated with the incremental inductance rule.

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3.4.3 Coplanar Waveguide with Ground

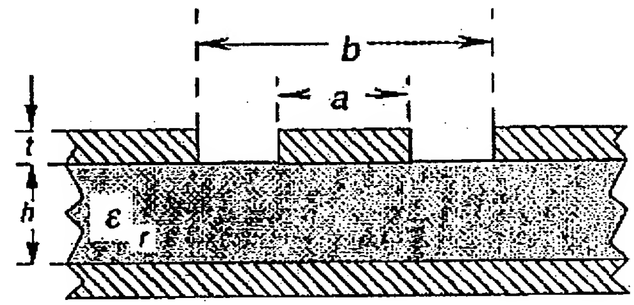

Figure 3.4.3.1: Coplanar Waveguide with Ground

The equations of this section can be used to analyze coplanar waveguide with ground or for microstrip lines with signal side ground plane.

$$Z_0 = \frac{60.0\,\pi}{\sqrt{\varepsilon_{eff}}} \frac{1.0}{\frac{K(k)}{K(k')} + \frac{K(k_1)}{K(k_1')}} \tag{3.4.3.1}$$

$$k = a/b \tag{3.4.3.2}$$

$$k' = \sqrt{1.0 - k^2} \tag{3.4.3.3}$$

$$k_1' = \sqrt{1.0 - k_1^2} \tag{3.4.3.4}$$

$$k_1 = \frac{\tanh\left(\frac{\pi a}{4.0\,h}\right)}{\tanh\left(\frac{\pi b}{4.0\,h}\right)} \tag{3.4.3.5}$$

$$\varepsilon_{eff} = \frac{1.0 + \varepsilon_r \frac{K(k')}{K(k)} \frac{K(k_1)}{K(k_1')}}{1.0 + \frac{K(k')}{K(k)} \frac{K(k_1)}{K(k_1')}} \tag{3.4.3.6}$$

The notation varies between references. Some use the a, b notation of Figure 3.4.2.1 or a $2a$ and $2b$ version while others use s (spacing to the adjacent ground) and w (trace width).

These equations "show good agreement" with spectral-domain variational calculation techniques.

The structure may propagate in three different modes: microstrip, coplanar waveguide, and coupled slotlines. To prevent slotline modes, jumpers connecting the two halves of the component side ground plane can be used. To avoid microstrip line modes, it is recommended [2] that $h >> b$ and that the component side ground extend away from the trace on each side more than b.

REFERENCES:

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3.4.4 Shielded Coplanar Waveguide

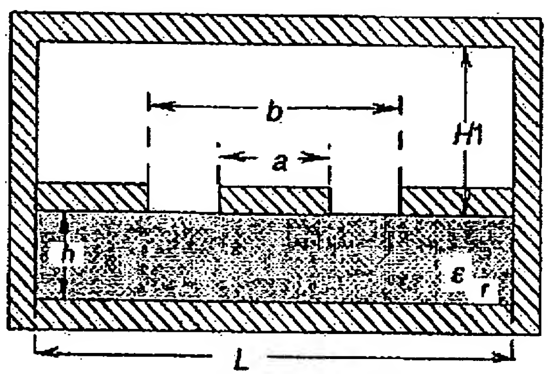

Figure 3.4.4.1: Shielded Coplanar Waveguide

The formulas below allow calculation of coplanar waveguide including the shield's effects [1, 2]. Many times we are simply looking for guidelines on how not to be affected by the shield's presence. For $L / b \geq 1.75$, and $H_1 / a \geq 2.50$, Z_0 is affected less than 1.5% by the shield's presence.

The equations of [2] assume H_1 is infinite and are:

$$Z_0 = \frac{1.0}{\dfrac{5.0\,q}{Z_m\,(1.0 + 5.0\,q)} + \dfrac{1.0}{Z_c(1.0 + q)}} \quad (\Omega) \qquad (3.4.4.1)$$

$$q = \frac{a}{h}\left(\frac{b}{a} - 1.0\right)\left[3.6 - 2\,e^{-(\varepsilon_r + 1.0)/4.0}\right] \qquad (3.4.4.2)$$

Z_c = coplanar waveguide impedance (3.4.4.3)

Z_m = microstrip impedance (3.4.4.4)

Use the equations given in the coplanar waveguide and microstrip line sections together with the relevant dimensions to calculate Z_c and Z_m. Accuracy is within 2% of numerical results.

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3.4.5 Asymmetric Coplanar Waveguide

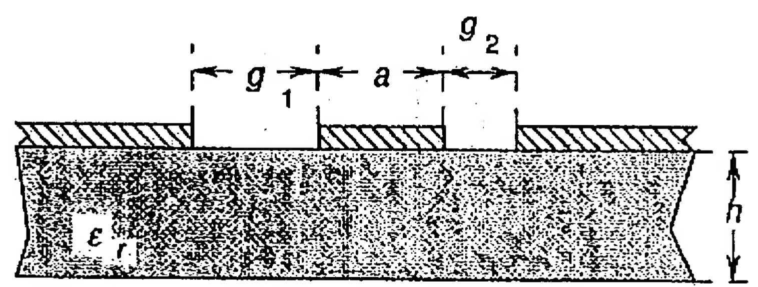

Figure 3.4.5.1: Asymmetric Coplanar Waveguide

The asymmetric coplanar waveguide structure can be used to reduce the line impedance of symmetric coplanar waveguide or to analyze the effects of fabrication tolerances.

The equations of Hanna and Thebault [1, 2] derived with conformal mapping are

$$Z_0 = \frac{30.0\,\pi}{\sqrt{\varepsilon_{eff}}} \frac{K'(k_1)}{K(k_1)} \quad (\Omega) \qquad (3.4.5.1)$$

$$\varepsilon_{eff} = 1.0 + \frac{\varepsilon_r - 1.0}{2.0} \frac{K(k_2)}{K'(k_2)} \frac{K'(k_1)}{K(k_1)} \qquad (3.4.5.2)$$

where

$$k_1 = \frac{0.5\,b\,[1.0 + \alpha\,(0.5\,b + d_1)]}{0.5\,b + d_1\,\alpha\,\sqrt{0.5\,b}} \qquad (3.4.5.3)$$

$$k_2 = \frac{w_A\,(1.0 + \alpha_1\,w_B)}{w_B + \alpha_1\,w_A^2} \qquad (3.4.5.4)$$

and

$$w_A = \sinh\left(\frac{\pi\,a}{4.0\,h}\right) \qquad (3.4.5.5)$$

$$w_B = \sinh\left[\frac{\pi\,(a\,/\,2.0 + g_1)}{2.0\,h}\right] \qquad (3.4.5.6)$$

$$w_E = -\sinh\left[\frac{\pi\,(a\,/\,2.0 + g_2)}{2.0\,h}\right] \qquad (3.4.5.7)$$

$$\alpha = \frac{d_1\,d_2 + 0.5\,b\,(d_1 + d_2) \pm \sqrt{d_1\,d_2\,(b + d_1)\,(b + d_2)}}{\sqrt{0.5\,b}\,(d_1 - d_2)} \qquad (3.4.5.8)$$

$$\alpha_1 = \left(\frac{1.0}{w_B + w_E}\right)\left[-1.0 - \frac{w_B w_E}{w_A^2} \pm \sqrt{\left(\frac{w_B^2}{w_A^2} - 1.0\right)\left(\frac{w_E^2}{w_A^2} - 1.0\right)}\,\right] \qquad (3.4.5.9)$$

The equations assume that the traces have negligible thickness. Comparison of the equations to experimental data showed agreement to within 4% for Z_0. Choose the '–' sign solution of Equations (3.4.5.8) and (3.4.5.9).

Graphs of the dispersion of this structure are available in [3].

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3.4.6 Coplanar Strips

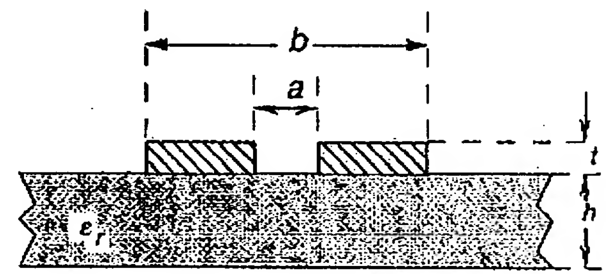

Figure 3.4.6.1: Coplanar Strips

The coplanar strips structure is similar to paired wire transmission line structures. Note also that coplanar strips are the complementary structure to coplanar waveguide.

$$Z_0 = \frac{\eta_0}{\sqrt{\varepsilon_{eff}}} \frac{K(k)}{K(k')} \qquad (3.4.6.1)$$

$$\varepsilon_{eff} = 1 + \frac{\varepsilon_r - 1}{2} \frac{K(k')\,K(k_1)}{K(k)\,K(k_1')} \qquad (3.4.6.2)$$

$$k = \frac{a}{b} \qquad (3.4.6.3)$$

$$k' = \sqrt{1.0 - k^2} \qquad (3.4.6.4)$$

$$k_1' = \sqrt{1.0 - k_1^2} \qquad (3.4.6.5)$$

$$k_1 = \frac{\sinh\left(\frac{\pi a}{4 h}\right)}{\sinh\left(\frac{\pi b}{4 h}\right)} \tag{3.4.6.6}$$

Losses for this structure are [4]

$$\alpha_c = 17.34 \frac{R_s}{Z_0} \frac{P'}{\pi a}\left(1.0 + \frac{b-a}{2.0 a}\right) \times \left(\frac{\frac{1.25}{\pi} \ln \frac{4.0 \pi (b-a)}{2.0 t} + 1.0 + \frac{2.5 t}{\pi (b-a)}}{\left\{1.0 + \frac{b-a}{a} + \frac{1.25 t}{\pi a}\left[1.0 + \ln \frac{2.0 \pi (b-a)}{t}\right]\right\}^2} \right) \quad (\text{dB / m}) \tag{3.4.6.7}$$

$$\alpha_d = \frac{20 \pi}{\ln (10)} \frac{\varepsilon_r}{\sqrt{\varepsilon_{eff}}} \frac{q \tan \delta}{\lambda_0} \quad (\text{dB / unit length}) \tag{3.4.6.8}$$

for $0 \le k \le 0.707$

$$P' = \frac{k}{k'^{3/2} (1.0 - k')} \left[\frac{K(k)}{K'(k)}\right] \tag{3.4.6.9}$$

for $0.707 \le k \le 1.0$

$$P' = \frac{1.0}{\sqrt{k}\,(1.0 - k)} \tag{3.4.6.10}$$

Pintzos [7] gives plots of the dispersion characteristics of this line.

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3.4.7 Asymmetrical Coplanar Strips

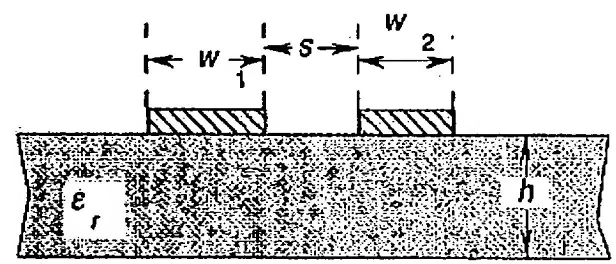

Figure 3.4.7.1: Asymmetrical Coplanar Strips

Hoffman [2] solves the equations for this structure from those of the infinitely thick dielectric structure. His equations are

$$Z_0 = \frac{\eta_0}{2.0 \sqrt{\varepsilon_{eff}}} \frac{K(k)}{K(k')} \quad (\Omega) \tag{3.4.7.1}$$

$$\varepsilon_{eff} = 1.0 - \frac{(\varepsilon_r - 1.0)\, K(k_1)\, K(k)}{2.0\, K(k_1')\, K(k')} \tag{3.4.7.2}$$

where

$$k = \sqrt{\frac{s}{k}\left(1.0 + \frac{b}{\cdot} - \frac{s}{\cdot}\right)} \tag{3.4.7.3}$$

$$k_1 = \sqrt{\frac{(t_1 - t_2)(t_3 - t_2)}{(t_1 + t_2)(t_3 + t_2)}} \qquad (3.4.7.4)$$

$$t_n = \frac{e^{\lambda_n} - 1.0}{e^{\lambda_n} + 1.0}, \quad n = 1, 2, 3 \qquad (3.4.7.5)$$

$$\lambda_1 = \frac{\pi}{2.0}\left(\frac{2.0\, w_2}{h} + \frac{s}{h}\right) \qquad (3.4.7.6)$$

$$\lambda_2 = \frac{\pi\, s}{2.0\, h} \qquad (3.4.7.7)$$

$$\lambda_3 = \frac{\pi}{2.0}\left(\frac{2.0\, w_1}{h} + \frac{s}{h}\right) \qquad (3.4.7.8)$$

$$k_1' = \sqrt{1.0 - k_1^2} \qquad (3.4.7.9)$$

$$k' = \sqrt{1.0 - k^2} \qquad (3.4.7.10)$$

$$b = w_2 + s \qquad (3.4.7.11)$$

$$d = w_1 + s \qquad (3.4.7.12)$$

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[1] Gevorgian, S.S., and I.G. Mironenko, "Asymmetric Coplanar-Strip Transmission Lines for MMIC and Integrated Optic Applications," *Electronic Letters*, Vol. 26, No. 22, October 25, 1990, pp. 1916–1918.

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3.4.8 Three Coplanar Strips

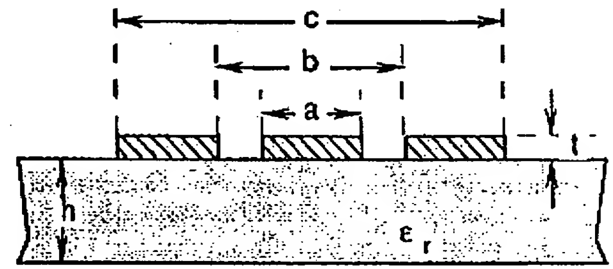

Figure 3.4.8.1: Three Coplanar Strips

This configuration was derived in [4] to analyze and design a tape automated bonding (TAB) IC package with alternate traces grounded.

$$Z_0 = \frac{\eta_0}{4.0\sqrt{\varepsilon_{eff}}}\frac{K(k_1)}{K(k_1')} \quad (\Omega) \qquad (3.4.8.1)$$

where

$$\varepsilon_{eff} = 1.0 + \frac{\varepsilon_r - 1.0}{2.0}\frac{K(k_2')}{K(k_2)}\frac{K(k_1)}{K(k_1')} + \frac{\varepsilon_r - 1.0}{2.0}\frac{K(k_2')}{K(k_2)}\left[\frac{K(k_1)}{K(k_1')}\right]^2\left(\frac{t}{b-a}\right) + \frac{2.0\, t}{b-a}\frac{K(k_1)}{K(k_1')} + \left[\frac{t}{b-a}\frac{K(k_1)}{K(k_1')}\right]^2 \qquad (3.4.8.2)$$

$$k_1 = \frac{c}{b}\sqrt{\frac{b^2 - a^2}{c^2 - a^2}} \qquad (3.4.8.3)$$

$$k_2 = \frac{\sinh\left(\frac{\pi\, c}{4.0\, h}\right)}{\sinh\left(\frac{\pi\, b}{4.0\, h}\right)}\sqrt{\frac{\sinh^2\left(\frac{\pi\, b}{4.0\, h}\right) - \sinh^2\left(\frac{\pi\, a}{4.0\, h}\right)}{\sinh^2\left(\frac{\pi\, c}{4.0\, h}\right) - \sinh^2\left(\frac{\pi\, a}{4.0\, h}\right)}} \qquad (3.4.8.4)$$

$$k_n' = \sqrt{1 - k_n^2} \quad n = 1, 2 \qquad (3.4.8.5)$$

In Houdart [3, Figure 2], the effect of ground plane width (c in Figure 3.4.8.1) on the CPW is analyzed. For $c/b > 5.0$ the impedance is affected by less than about 3%.

REFERENCES

[1] Ghione, Giovanni, and Carlo U. Naldi, "Coplanar Waveguides for MMIC Applications: Effect of Upper Shielding Conductor Backing, Finite-Extent Ground Planes, and Line-to-Line Coupling," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-35, No. 3, March 1987, pp. 260–267.

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3.4.9 Three Coplanar Strips with Ground

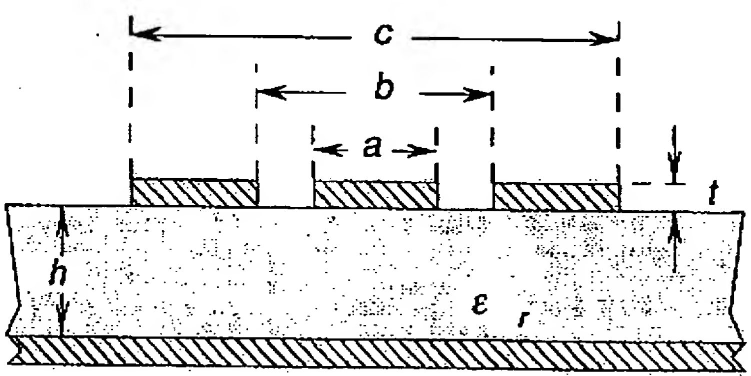

Figure 3.4.9.1: Three Coplanar Strips With Ground

This configuration was derived in [1] to analyze and design a tape automated bonding structure where alternate traces are grounded and are connected to a ground layer with vias.

$$Z_0 = \frac{\eta_0}{2.0\sqrt{\varepsilon_{eff}}} \frac{1.0}{\frac{K(k_1)}{K(k_1')} + \frac{K(k_2)}{K(k_2')} + \frac{2.0\,t}{(b-a)}} \quad (3.4.9.1)$$

where

$$\varepsilon_{eff} = \frac{\frac{K(k_1)}{K(k_1')} + \frac{\varepsilon_r K(k_2)}{K(k_2')} + \frac{2.0\,t}{b-a}}{\frac{K(k_1)}{K(k_1')} + \frac{K(k_2)}{K(k_2')} + \frac{2.0\,t}{(b-a)}} \quad (3.4.9.2)$$

$$k_1 = \frac{c}{b}\sqrt{\frac{b^2-a^2}{c^2-a^2}} \quad (3.4.9.3)$$

$$k_2 = \frac{\tanh\left(\frac{\pi\, a}{4.0\, h}\right)}{\tanh\left(\frac{\pi\, b}{4.0\, h}\right)} \quad (3.4.9.4)$$

$$k_n' = \sqrt{1.0 - k_n^2}\,,\ n = 1, 2 \quad (3.4.9.5)$$

REFERENCES

[1] Wentworth, Stuart M., *et al.*, "The High-Frequency Characteristics of Tape Automated Bonding (TAB) Interconnects," *IEEE Transactions on Components, Hybrids, and Manufacturing Technology*, Vol. CHMT-122, No. 3, September 1989, pp. 340–347.

3.4.10 Covered Coplanar Waveguide with Ground

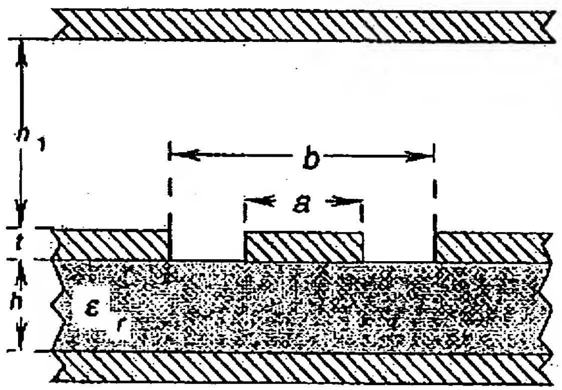

Figure 3.4.10: Covered Coplanar Waveguide with Ground

The equations describing this structure are

$$Z_0 = \frac{\eta_0}{2.0\sqrt{\varepsilon_{eff}}} \frac{1.0}{\frac{K(k_3)}{K(k'_3)} + \frac{K(k_4)}{K(k'_4)}} \quad (3.4.10.1)$$

$$\varepsilon_{eff} = 1.0 + \frac{\frac{K(k_3)}{K(k'_3)}}{\frac{K(k_3)}{K(k'_3)} + \frac{K(k_4)}{K(k'_4)}} (\varepsilon_r - 1.0) \quad (3.4.10.2)$$

where

$$k_3 = \frac{\tanh\left(\frac{\pi a}{h}\right)}{\tanh\left(\frac{\pi b}{h}\right)} \quad (3.4.10.3)$$

$$k_4 = \frac{\tanh\left(\frac{\pi a}{h_1}\right)}{\tanh\left(\frac{\pi b}{h_1}\right)} \quad (3.4.10.4)$$

which are strictly valid for

and were found to give good results elsewhere.

REFERENCES

[1] Ghione, Giovanni, and Carlo U. Naldi, "Coplanar Waveguides for MMIC Applications: Effect of Upper Shielding Conductor Backing, Finite-Extent Ground Planes, and Line-to-Line Coupling," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-35, No. 3, March 1987, pp. 260–267.

3.4.11 Covered Coplanar Waveguide

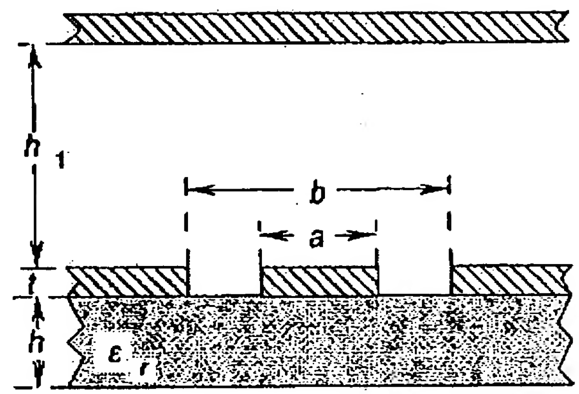

Figure 3.4.11.1: Covered Coplanar Waveguide

The equations of Ghione and Naldi [1] are

$$Z_0 = \frac{\eta_0}{2.0\sqrt{\varepsilon_{eff}}} \frac{1.0}{\frac{K(k_2)}{K(k'_2)} + \frac{K(k)}{K(k')}} \quad (3.4.11.1)$$

$$\varepsilon_{eff} = 1.0 + \frac{\frac{K(k_1)}{K(k'_1)}}{\frac{K(k_2)}{K(k'_2)} + \frac{K(k)}{K(k')}} (\varepsilon_r - 1.0) \quad (3.4.11.2)$$

where

$$k = a / b \quad (3.4.11.3)$$

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Drawings

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Receipt is hereby acknowledged of the papers filed as indicated in connection with the above-identified case.   MRQ/YAP